Some experiments with high-frequency ladder crystal filters

by J. A. HARDCASTLE, G3JIR*

Part 1. Construction

Introduction

Although electrical networks textbooks have long acknowledged the existence of this type of filter, almost nothing has been written about them in amateur radio journals. It is hoped that this article will rectify the situation and will show how simply an effective filter may be produced using only a handful of crystals and capacitors. The crystals used are of identical frequency, while the capacitors are silvered mica two per cent tolerance and preset trimmer types. An experimental approach has been adopted throughout and full details of test procedures and the results obtained are given.

Half-lattice filters

Early attempts by the author to make high frequency crystal filters using 8·3MHz 10X and FT243 type surplus crystals in the familiar half-lattice, four-crystal configuration (Fig 1), produced such poor results that they never progressed beyond the breadboard test stage. They suffered from too narrow a bandwidth (2kHz), poor stop-band discrimination (40dB) and numerous spurious responses immediately adjacent to the hf cut-off frequency. It was considered that while they would be suitable for an ssb upper-sideband filter for a transmitter, they would be unusable in a receiver.

Ladder filters

Ladder-type filters, however, are not so vulnerable to the effects of the additional series resonances present in the crystals. This is because of the unlikelihood of these resonances occurring at identical frequencies in all crystals, so the resonances of one section are attenuated by all the other sections.

Armed with this knowledge, the author sought a cheap source of suitable crystals. Happily HC6U crystals were

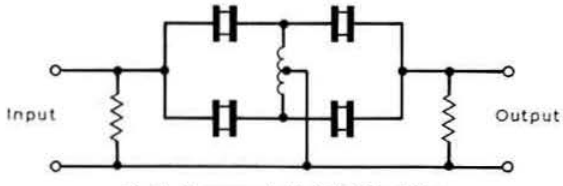

Fig 1. Four-crystal half-lattice filter

*82 Acacia Avenue, Huyton, Liverpool L36 5TP.

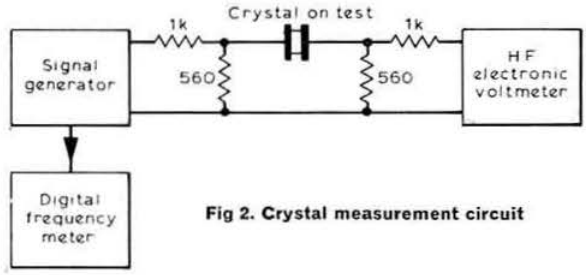

Fig 2. Crystal measurement circuit

being advertised in *Radio Communication* at an attractive price and the advertiser readily agreed to supply a batch of his choice of frequency, between 9 and 10MHz.

Crystal measurements

The width of the passband of any crystal filter is dependent on the spacing of the series resonant frequency and the parallel resonant frequency of the crystal. The test circuit shown in Fig 2 was made so that these frequencies could be checked, and comprises a signal generator, a digital frequency meter and a sensitive hf electronic voltmeter. A typical test measurement made with this equipment is shown in Fig 3(a). In order to demonstrate the presence of numerous series and parallel resonances the response has been drawn out in full, but it is generally sufficient to check only the first pair of these frequencies. Table 1 summarizes measurements made on a number of different types of crystal and it is immediately apparent that the plated type of crystal is capable of a much wider bandwidth than the older type of clamped mounting. Clearly the 9·6MHz HC6U crystals will readily achieve the 2·4kHz bandwidth required for an ssb filter.

First steps

Before attempting to produce a high-performance, multiple-section ladder filter it was decided to add two shunt capacitors to the test circuit of Fig 2, transforming it into a simple bandpass filter (Fig 4).

Using various values of capacitance for C1 and C2, the frequency responses were measured so that some idea could be obtained of the size of the components which would be required in a full-sized filter. The result of one of these tests is shown in Fig 3(b) where C1 and C2 were each of 150pF.

An intermediate stage

Expanding the simple filter of Fig 4 to three sections proved to be an easy operation. When three identical filter sections (Fig 5(a)) are combined, the resulting circuit becomes that

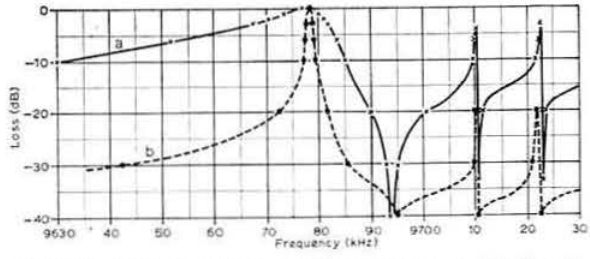

Fig 3.(a) Crystal parallel and series resonances. (b) Simple bandpass filter

Table 1. Crystal measurements

Nominal frequency f_o (kHz)	Case Reference	Mounting	Series resonant frequency f_s (kHz)	Parallel resonant frequency f_p (kHz)	$f_p - f_s$ (kHz)	$\frac{f_p - f_s}{f_o}$ %
8,300	FT243	Compression	8,310·99	8,312·45	1·46	0·018
8,300	CR1AR	Compression	8,306·09	8,308·2	2·11	0·025
8,300	10X	Compression	8,309·7	8,312·52	2·82	0·034
8,325	10XJ	Plated	8,324·05	8,328·86	4·81	0·057
3,250	HC6U	Plated	3,250·8	3,254·7	3·9	0·12
40,500 (13,500)	HC6U	Plated (overtone)	13,507·6	13,524·1	16·5	0·12
9,681·2	HC6U	Plated	9,672·55	9,687·95	15·4	0·16

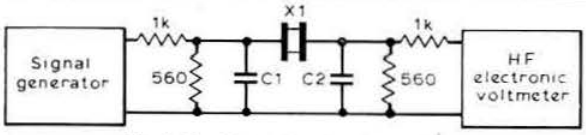

Fig 4. Single-section bandpass filter

of Fig 5(b). Note how connecting similar stages in tandem results in the end-section capacitor being half the value of the others.

A quick scan over the frequency band showed that the circuit was behaving like a bandpass filter, with a sharp cut-off on each side of the pass-band. However, the shape of the passband left a great deal to be desired and the test set-up was insufficiently sensitive to enable adequate stop-band measurements to be made.

In order to solve the sensitivity problem a simple superhet receiver, shown in broad outline in Fig 6, was constructed; this will be described in detail in Part 2. All the tests described subsequently were made using this test set.

The shape of the passband was improved by providing the correct source and load impedance in the form of preset variable resistors, as shown in Fig 7. The frequency characteristics of these two filters (Fig 8) show how careful adjustment of R1 and R2 has resulted in a passband ripple of less than 1·5dB. R1 is a shunt resistor to reduce the output impedance of the buffer amplifier and R2 is a series resistor to increase the input impedance of the rf amplifier. When the filter is used in a circuit of the correct impedance, R1 and R2 will no longer be required, but for the purpose of these tests they enabled a variety of filters to be tested quickly, without the necessity of continually modifying the test equipment.

Note the fringe of spurious responses which were found above 9,710kHz. More than six of these extremely sharp responses were found—they are only about 100Hz wide—and reference back to Fig 3(a) shows how they correspond with two of those shown there. However, as mentioned previously, they have been greatly attenuated by the other two stages.

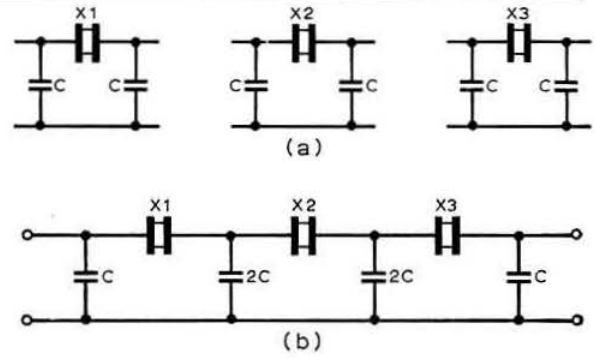

Fig 5. Addition of three single-filter sections to produce a three-section filter

Bandwidth

It was found that the bandwidth of the three-section filter could be controlled by selection of the size of the shunt capacitors. By increasing their capacity the bandwidth could be reduced, but this also reduces the filter's impedance and so necessitates the readjustment of R1 and R2. Figs 7(b) and 8(b) show the result of a 50 per cent increase in capacity, which reduced the 3dB bandwidth from 3,500Hz to 2,600Hz. It was impossible to achieve a completely satisfactory passband ripple by adjusting R1 and R2 alone, but when C4 was reduced to 50pF a ripple of 1dB was obtained. Note also how the stop-bands have been affected by the change, the lf being improved and the hf degraded.

At this point it was considered that sufficient experience of the behaviour of simple filters had been acquired and that an attempt to produce a full-sized filter should be made.

Six-section filter

Adding three more crystals produced the filter of Fig 9, the values used for the capacitors being derived from the three-section filter described previously. As expected from previous measurements, stop-bands of greater than 70dB

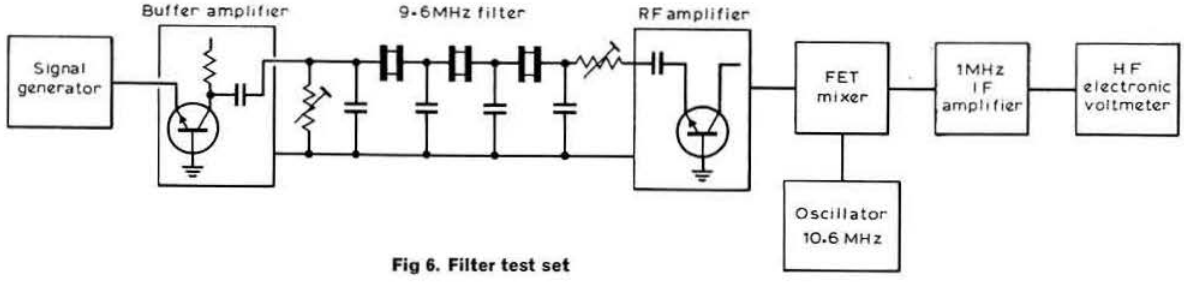

Fig 6. Filter test set

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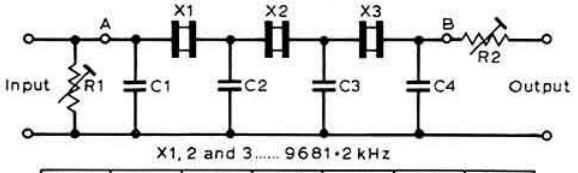

Fig	R1	C1	C2	C3	C4	R2
7a	3 kΩ	50pF	100pF	100pF	50pF	330Ω
7b	1·8 kΩ	75pF	150pF	150pF	50pF	200Ω

Fig 7. Three-section filters

were easily produced and the region requiring most effort was again the passband. Besides adjusting R1 and R2 it was also found necessary to adjust C1 and C7. In fact C7 was eventually removed altogether, stray capacitance alone being sufficient in this position. As a final contribution to a flat pass-band, C2 and C6 were also slightly reduced and the resulting ripple in the response is 1dB.

The bandwidth at −3dB is 2,757Hz, at −6dB 2,923Hz and at −60dB 6,698Hz, giving a 60:6dB shape factor of 2·29:1. The insertion loss, measured between points A and B in Fig 9, is 3dB. Fig 10 shows the full frequency response.

Filters with the same capacitor values, using five and seven crystals, were also checked. The characteristics of these filters have not been given here, but the five-crystal filter had a shape factor of 2·93:1 and the seven-crystal filter a shape factor of 1·89:1. In both cases the 6dB bandwidth was almost identical with the six-crystal filter. The response, at −60dB only, is plotted in Fig 10 for ease of comparison.

Conclusion

These experiments with high-frequency ladder filters have led the author to conclude that they have several advantages over lattice-type filters, particularly for the amateur. They may be summarized as follows:

(1) All crystals are of the same frequency and no matching, grinding or etching is required.

(2) Spurious responses are not so detrimental to the overall performance of the filter and, for filters having more than four sections, may be virtually undetectable.

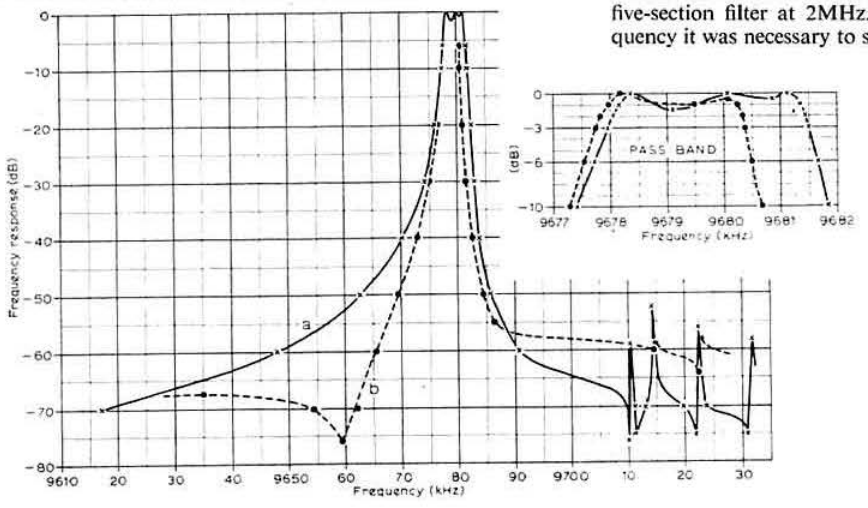

Fig 8. Three-section crystal ladder filters

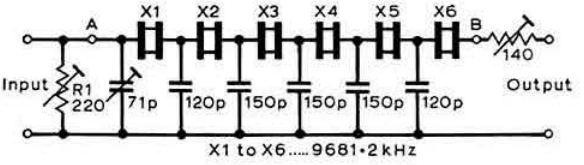

Fig 9. Six-section crystal ladder filter

(3) Filters may be constructed using an odd or even number of crystals; this is a very useful attribute when one is dependent on surplus sources for the supply of crystals.

(4) The only other components required are two per cent tolerance silvered mica capacitors and preset trimmers, which are readily available.

(5) Due to the very low equivalent series resistance of modern crystals the insertion loss of these filters is very low.

Job lots of suitable HC6U crystals have been offered by advertisers in *Radio Communication*, usually as batches of "our choice of frequency". Perhaps these will now become available as lots of one frequency; or an enterprising crystal manufacturer may decide to mass-produce 9MHz crystals at a price attractive to amateurs. Finally, when buying groups of crystals remember to buy an extra one for use as a carrier crystal.

Bibliography

As mentioned earlier not a great deal has been published in amateur magazines about crystal ladder filters, but two articles were found by the author which clearly pointed the direction for the present series of experiments.

[1] "An introduction to crystal filters", Hamish V. Bell, G3MAZ, *RSGB Bulletin* January-February 1962.

Part 2 of this general review article contained a description of a simple filter using an FT243 crystal in a circuit which was very similar to Fig 4. G3MAZ noted the shortcomings of these crystals for use in filters, but clearly recognized the usefulness of this design when suitable crystals became available.

[2] "PE wideband hf communications receiver", R. Hirst, *Practical Electronics* February 1970.

Part 5 of this series discussed the ssb filter which was a five-section filter at 2MHz. Due to the relatively low frequency it was necessary to select crystals which were slightly

Continued on p905

Components list (cont'd)

X1	Crystal 11·155
X2	Crystal 80·5MHz
Mixer	Hatfield 1759
F1, 2	10·7MHz 250kHz Toko CFS or Murata
F3, 5	455 8kHz Toko CFT 455B
F4	455 6kHz Toko CFT 455C
F6	455 4kHz Toko MFH41T
L1	7t on ¼in dia tapped at 3 & 4t 22swg
L2	2½t ¾in dia on wiring side of pcb
L3, 4	420μH on LA1157 100t 38swg wire
L5	24t 32swg on ⅞in neosid former with tuning slug but no screening can
L6	6t on ¼in dia 22swg tinned copper
L7	10t on ¼in dia 26swg enamelled
L8, 9	10μH Plessey choke
L10	100μH
T1, 2	6t 22swg tinned on ¼in dia + 1t link pvc covered wire (2x) 22swg
T3	FX2073 pri 8 + 8t sec 8t 32swg
T4	10t on ¼in dia 26swg enamelled + 1t 22swg pvc covered
T5	6t on ⅞in dia 22swg tinned copper + 1t 22swg pvc covered (2x)
T6	Toroid FX2073 pri 20t sec 10 + 10t 32swg enamelled
T7	6t on ¼in dia 22swg tinned copper + 2t 22swg pvc

out in this respect. Switching out one gang of the variable capacitor and padding correctly produces a second range from 144-145MHz.

The second point simply involves eliminating the inconvenience of tuning through the band when switching on in order to lock the phase locked loop.

Third, the 8kHz-wide filter is too narrow for fm and causes some distortion. It would probably help if the two 8kHz filters were replaced with i.f. transformers. The filter does, however, pull fm signals out of the noise extremely well and this narrowness has little effect on weaker signals.

These points have not been raised because the author feels disappointed with the results, but so that others may benefit from them, as he is extremely pleased with the way the receiver performs. A check with a signal generator showed that 1μV was a considerable signal on the meter compared with readable signals from an aerial. Stability is excellent and the whole apparatus takes 0·25A from a 12V battery, with the dial lights taking another 0·25A. □

Some experiments with hf ladder crystal filters

Continued from p898

staggered in frequency. The graph of the filter's performance showed a passband ripple of 5dB which is rather excessive.

On a professional level the following paper by a well-known writer on half-lattice hf crystal filters was the first detailed article on hf ladder filters to come to the author's attention:

[3] "A new approach to the design of high frequency crystal filters", R. A. Sykes. *Bell System Monograph 3180.*
[4] *Handbook of filter synthesis*, A. I. Zverev.

This is a reference work for the specialist, giving design information on all types of filter. Chapter 8 on crystal filters

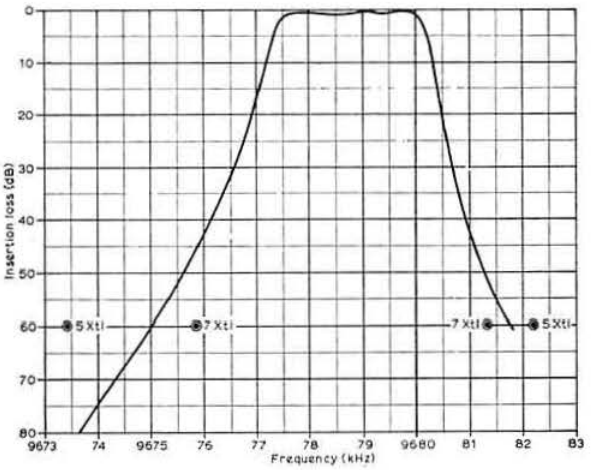

Fig 10. Six-section filter response. Response of five- and seven-section filters is also indicated

gives considerable information on the design of hf crystal filters.

Postscript

As this article was nearing completion the September 1976 issue of *Wireless World* came to hand, to be closely followed by the September issue of *Radio Communication*. The description by G3VA of the designs of F6BQP contained therein have been read with great interest and it is hoped to be able to test them in the near future for comparison with the designs which have been described here.

Part 2, Test equipment next month. □

Low pass filters

Continued from p899

amateur bands, the impedance/phase angle and the standing wave ratio. The table gives the attenuation in decibels for four specified frequency bands corresponding to Swedish television channels in Band 1 together with the attenuation above 70MHz.

Attenuation in decibels for four bands of frequencies and above 70MHz

Filter	Ch 1 33·4–38·9	Ch 2 48·25–53·75	Ch 3 55·25–60·75	Ch 4 62·25–67·75	>70MHz
TV-1000-LP	0·5– 1·5	2·5– 12	15 – 55	66–80	55
TV-3300-LP	1·8– 55	72 –100	77 –100	75–77	80
Johnson Viking	0·8	20 – 40	45 – 60	100	80
Trio LP30	0·5– 3	43 – 50	51 – 58	65–75	95
Ameco LN2	3 – 6	9 – 11	11·5– 13	14–16	20
Home-built	20 – 60	80 – 85	88 – 90	92–95	100
Trio LF-30A	6 – 30	45 –58	60 – 70	71–80	80
Yaesu FF-50DX	0·5– 20	50 – 62	50 – 52	53–58	60

In conclusion, the author gives a performance figure for each of the commercial filters. This figure lies between 0 and a maximum of 5.

Drake TV-3300-LP	4	Yaesu FF-50DX	2·5
Johnson Viking	4	Trio LP30	2
Trio LF-30A	3·5	Ameco LN2	0
Drake TV-1000-LP	3		

G2BVN □

Some experiments with high-frequency ladder crystal filters

by J. A. HARDCASTLE, G3JIR*

Part 2. Test equipment

Introduction

In order to be able to report some meaningful results it was necessary that the measurements made on the experimental filters should be highly accurate, both in respect of frequency and of insertion loss, and the author was fortunate in having access to a digital frequency meter and a sensitive rf voltmeter, as well as a suitable signal generator. Even so, the performance of the rf voltmeter was inadequate in respect of sensitivity at 9·6MHz, and the test set to be described was made to overcome this deficiency. However, the potential constructor should not be put off by this expensive array of test equipment, because it is possible to substitute a vfo or vxo for the signal generator and a station receiver for the rf voltmeter.

Insertion loss test set

The general arrangement of the equipment for this test is shown in Fig 11, the section between A and B being a simple superhet receiver specially designed for the measurement of the stopbands of crystal filters. The full circuit is shown in Fig 12 and is described section by section.

Buffer amplifier

TR1 is a grounded base amplifier and provides an impedance transformation from the 50Ω output of the signal generator to the 250-750Ω impedance range of the filters being tested. The collector circuit is broadly tuned to 9·6MHz by L1 and its output impedance is approximately 1·8kΩ. This stage is designed to handle the high level of signal required for measuring filter stopbands, and an output of 1·4V (4V peak to peak) into a 150Ω load is available for 0·4V input—a level which most signal generators are capable of supplying.

*82 Acacia Avenue, Huyton, Liverpool L36 5TP.

RF amplifier

Following the filter is TR2, another grounded base stage similar to the buffer amplifier, and also broadly tuned to the filter frequency. The input impedance is very low, approximately 10Ω, hence the need for a series matching resistor for the filter. Grounded base stages were chosen for these first two stages for their ability to provide rf amplification without instability problems over a wide range of source and load impedances.

Mixer

A fet mixer, TR3, is used to convert the input frequency to an i.f. of 1MHz. Because the test set was planned to be capable of use at any frequency in the hf range no internal oscillator was provided. A crystal oscillator, a vfo or yet another signal generator may be used according to availability. There is only one tuned circuit at 1MHz because in this case only limited selectivity is required in order to prevent the selectivity of the filter under test being enhanced by the test set.

The local oscillator level is set at 120mV rms, the minimum level at which the mixer will operate correctly. This level is usually specified as half the pinch-off voltage of the fet, but the 2N3819 which was used required −2·6V for cut-off and an oscillator level of 0·42V (1·3V peak to peak) caused too high a level of 10·6MHz signal to appear at the output. This is a consequence of having only one tuned circuit at the i.f.; however, a lower oscillator level and a 10·6MHz rejection circuit combine effectively to remove this unwanted signal.

Further gain at the i.f. is provided by TR4, which will give an output of 2·5V (7V peak to peak), into a 2kΩ load. The relatively low output impedance of this stage allows the use of a couple of feet of coaxial cable to connect to the rf voltmeter without too much reduction of the gain of the stage. An emitter follower could have been used to match to the cable but it would have been an additional stage and, more importantly, emitter followers have a tendency to oscillate when feeding a capacitive load.

Performance

The bandwidth of the test set is 50kHz at −3dB, and 85kHz at −6dB, which is sufficiently wide not to affect measurements on a narrow-band filter.

Measurements may be made over the full range of a bandpass filter from one stopband to the other, without changing the signal generator output level attenuator. This is a result of the test set being designed to handle a wide range of signal levels. It has proved to be of considerable operating convenience, because the signal generator frequency was found to shift when the output attenuator was adjusted. This produced disconcerting effects when making

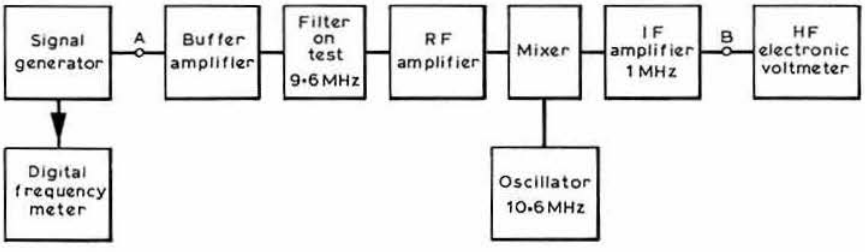

Fig 11. General arrangement of filter test set

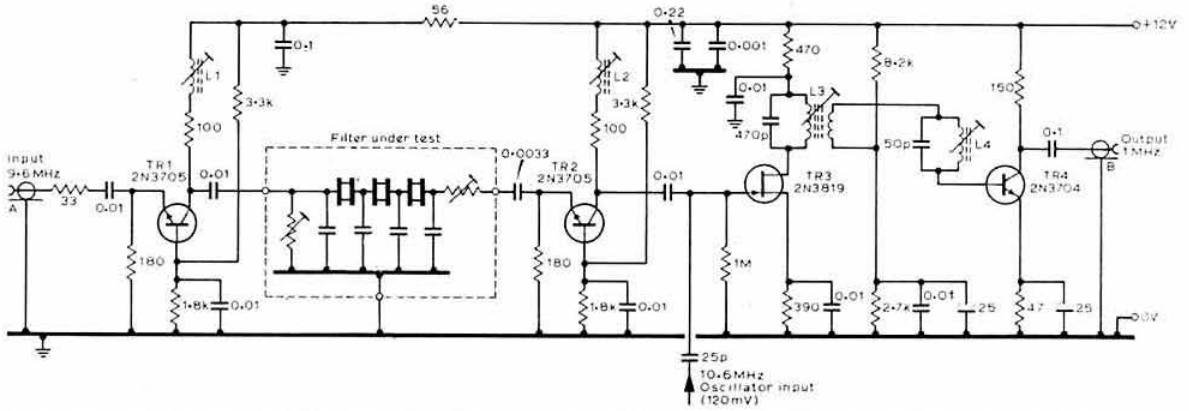

Fig 12. Circuit details of filter test set. Coil details: L1, 30t; L2, 30t; L3, primary, 62t, secondary, 12t; L4 33t; all 36swg enamelled copper wire on ¼in diameter former with iron dust core

measurements near the edge of the passband, a small adjustment of the attenuator causing the signal to disappear altogether! Now the output may be set to a level required to measure the stopbands and no further adjustments need be made to it as the frequency characteristic is measured. As the output level from the test set falls as the signal is tuned from the passband to the stopband, the rf voltmeter is switched to a more sensitive range. Having only one range switch to change considerably speeds up the measurements.

RF voltmeters

A number of rf voltmeter designs have been published [1, 2, 3]. The main requirements for the purposes of this article are a sensitivity range of 1mV-2·5V and a frequency range up to at least 1MHz.

Receiver as detector

As mentioned earlier a receiver may be used as a detector and if a switched attenuator is available a set of readings may be obtained almost as readily as when using an rf voltmeter. Fig 13 shows how the rf voltmeter has been replaced by the switched attenuator and receiver.

Tune the vfo to a frequency in the stopband and, without the filter in circuit, tune the receiver to this frequency. Now insert the filter and adjust the receiver sensitivity to give a mid-scale S-meter reading, with the attenuator set to zero. Then tune the vfo toward the passband and retune the receiver to peak up the signal. As the passband is approached it will be necessary to increase the attenuator setting in order to maintain the S-meter reading constant. Note the attenuator reading and the frequency at each point; as before, this will enable an attenuation characteristic to be plotted for the filter. Again the need for continually retuning the receiver must be stressed, because failure to do this will result in the selectivity characteristic of the receiver being plotted, rather than the filter under test!

Fig 13. Filter measurement using receiver

References

[1] "Silicon transistor millivoltmeter", D. E. O'N. Waddington, *Wireless World*, March 1966.
[2] "Linear scale millivoltmeter", A. J. Ewins, *Wireless World*, December 1970.
[3] "*Wireless World* ac millivoltmeter", *Wireless World*, June 1964. □

The Smith Chart

(Continued from p27)

$+1{\cdot}5j$ in series, ie $900j\Omega$ which is an inductance of 10·6μH. If the idea of a pair of large coils which need weather protection is disliked, continue a further 0·35λ towards the transmitter, once more crossing the 1·0 resistance circle point J. This time the reactance is $+1{\cdot}5j$ and requires a series tuning capacitance of 12·7pF, eg a pair of 25pF capacitors in series, one in each wire so that the line remains balanced.

More often than not, in cases such as the last example it is preferable to connect matching components or stubs in parallel with the line. The discussion so far has given no indication of one of the most valuable features of the chart, namely the ease with which parallel/series substitutions can be made—assuming familiarity with admittance, which is the reciprocal of impedance, and use of inverting properties of the chart to effect this type of transformation (to be explained in the next article).

(6) *An antenna for 14MHz has an impedance of 40Ω, providing an acceptable match to 50Ω cable. Unfortunately only 95ft of 50Ω feeder is available and the total length required is 106ft. What will be the effect on swr if λ/4 of 75Ω feeder is used to make up the required length? Where is the best place to put it? Should it be at the transmitting or the antenna end of the 50Ω line, or somewhere in between? More 75Ω feeder is available if it can be used to advantage, and the velocity factor for both cables is 0·67.*

The electrical length of the 50Ω feeder is 2·125λ, which from Fig 8, goes from the antenna, point K to a point M on the chart, an impedance of $0{\cdot}97 + 0{\cdot}22j$. Adding on the λ/4 of 75Ω cable this value must be normalized for the new impedance, multiplying by 50 and dividing by 75, ie multiplying by two-thirds to give an impedance of $0{\cdot}65 + 0{\cdot}15j$. This is point M′ and fits on to a swr circle of 1·6 which is the value applicable to the 75Ω line. If it is desired to know the impedance at the transmitting end of the line, this can be found by proceeding λ/4 round the 1·6 swr circle to the point N.

Next suppose the extra length is placed at the antenna end of the line. The starting point is now 40/75 but is inverted by the λ/4 line to become 75/40 at the start of the 50Ω cable. Normalizing to 50Ω involves a multiplying factor of 1·5, bringing the resistance and therefore the swr up to 2·74 which is much higher than before and applies to most of the feeder run, whereas the previous highest value 1·6 applied only to the last 11ft.

Much time can be spent in trying to improve on the first result, since the chart allows experimentation on paper with a wide range of cable lengths and positions, but the following is a good solution. Proceeding from the antenna at 50Ω for λ/8 and then changing to 75Ω again gives point M′. After a distance of 0·476λ the point R is reached. This is a resistance of 0·63, ie 47·5Ω and re-normalizing to 50Ω this becomes 0·95, providing an swr of 1·05 in the 50Ω cable which may be used for the remaining 90ft to the transmitter. □

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by J. A. HARDCASTLE, G3JIR*

Part 3. Filters for transmitters and receivers

Introduction

In this part, ssb transmitter filters are described and a set of capacitor coefficients for a six-crystal filter is given, and from this a seven-crystal, symmetrical receiver filter is derived. Before describing these new developments some test results on F6BQP filters [1] are reported.

F6BQP design

Three- and four-crystal filters (Fig 14) were constructed using the same crystals as before, so the effect of changing capacitor values and circuit impedances could be compared. The components used were the nearest standard values to those calculated from F6BQP's formulas. Fig 15 compares the performance of F6BQP and G3JIR three-crystal filters, showing how F6BQP has given a filter design having a smooth passband (Butterworth filter), while the G3JIR filter has a slightly better rate of cut-off at the expense of some passband ripple (Chebyshev filter). Exchange of skirt selectivity for passband ripple is a familiar feature of filter design commonly encountered in LC filters. The degree of ripple tolerable is a subjective decision and has not been investigated during these experiments.

A graph of the four-crystal filter response is not included because it is very similar in shape to the three-crystal filter. It is 3dB down at 9,677·8kHz and 9,680·4kHz, and 50dB down at 9,670·2kHz and 9,683·5kHz.

Lower sideband filter

A filter having few crystals has a distinctly asymmetric frequency response (Fig 15) which can be accentuated by

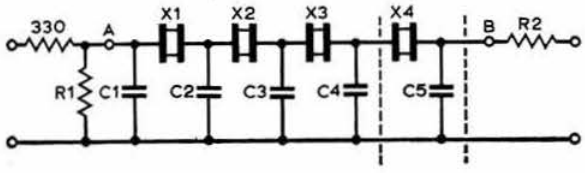


Fig 14. Three- and four-crystal filters.

	No of crystals	Filter type	R1 Ω	R2 Ω	C1	C2	C3 pF	C4	C5
(a)	3	F6BQP	230	230	50	150	150	50	—
(b)	3	G3JIR	350	350	75	150	150	75	—
(c)	4	F6BQP	310	310	25	100	150	100	25

*82 Acacia Avenue, Huyton, Liverpool L36 5TP.

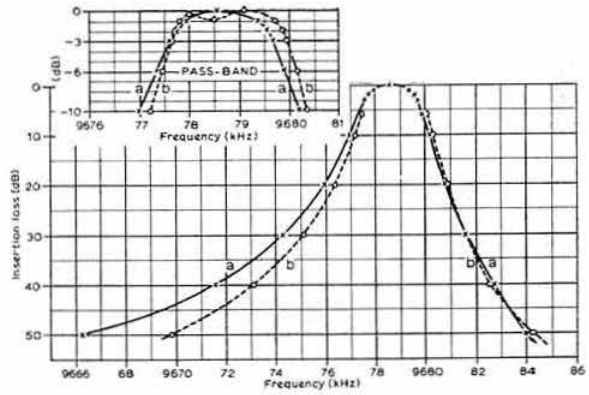


Fig 15. Three-crystal filters. (a) F6BQP; (b) G3JIR

shunting some, or all, of the crystals with additional capacitance. This reduces their parallel resonant frequency, bringing it closer to the passband and increasing the slope of the hf skirt. In Fig 16(a) three of the crystals are shunted with small trimmer capacitors adjusted to make the hf skirt as steep as possible, while maintaining almost 60dB stopband rejection.

Fig 17(b) illustrates the need for spacing out rejection notches to achieve a uniform stopband. If rejection notches are brought closer to the band edge, by further increasing shunt capacitance, greater rejection is obtained over a limited band, after which it steadily deteriorates. This added capacitance causes some deterioration of lf skirt selectivity, but like remote regions of the hf stopband this is not so important in an ssb transmitter, as in a receiver; particularly if audio bandwidth is limited before the modulator to reduce the width of generated sidebands.

Upper sideband filter

An upper sideband filter may be made as shown in Fig 16(b). The passband is now controlled by the crystal's parallel resonant frequency and the stopband by its series resonant frequency. Because all of these crystals are of the same frequency a single deep rejection notch is produced at 9,678kHz, after which the stopband deteriorates. The passband in Fig 18(a) is 7kHz wide but in (b) it has been reduced to 4kHz by adding shunt capacitors to three of the crystals.

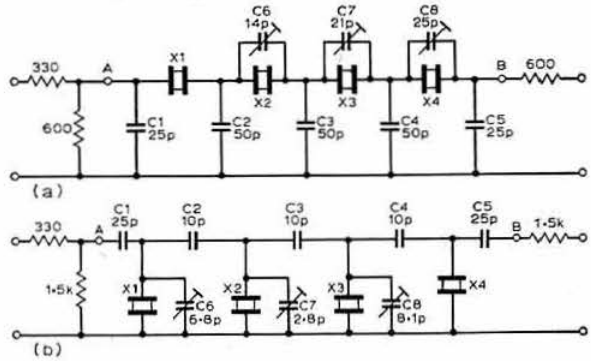


Fig 16. SSB filters. (a) Lower sideband; (b) Upper sideband; X1-X4 9,681·2kHz

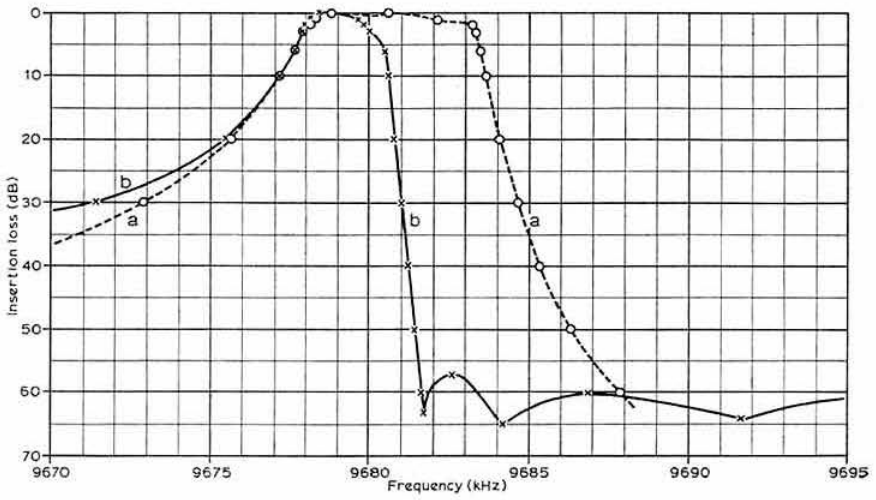


Fig 17. Lower sideband filter. (a) Without C6, 7, 8; (b) With C6, 7, 8

Upper sideband filters need to operate in a high impedance circuit and great care has to be taken over screening to ensure a good performance. While testing these filters it was noticed that the buffet amplifier output circuit was slightly influencing the passband shape, so a 330Ω resistor was connected in series with the input, to provide some isolation.

Dishal filters

Since writing the two earlier parts of this article an important paper [2] has been brought to the author's attention, by courtesy of G4CUE and G3VA. It describes designs for upper- and lower-sideband filters, and gives tables of design coefficients for Butterworth and Chebyshev filters.

Capacitor coefficients (Fig 19) for the six-crystal Chebyshev design have been derived from this information. These allow a filter to be designed without knowledge of the crystal's parameters, the only penalty being an inability to calculate the filter bandwidth, and this must be found by experiment; however, Figs 19 and 20 will provide a starting point for future designs. Butterworth filters designed from the data given in [2] were also constructed and tested successfully.

In both designs components calculated from these coefficients gave the required response shape without further adjustment. Dishal gives more design coefficients in [3] from which a wide range of filters may be designed.

Seven-crystal filter

A symmetrical filter for a receiver may be made by adding a shunt crystal (X7) to the Dishal filter (Fig 19). This is of the same nominal frequency as X1 to X6 but its series resonant frequency has been reduced by adding inductance L1. This combination produces a rejection notch which can be tuned by L1 to obtain the optimum If skirt response (Fig 20(b)). Reducing the inductance increases the notch frequency, steepening the If skirt, but if carried to excess this will degrade the If stopband. Adding X7 increases the passband ripple to 2dB but has no effect on the hf stopband.

Carrier oscillator

The series mode carrier oscillator circuit (Fig 21) was found to have useful vxo properties. Crystals can be pulled over the range 9,676kHz to 9,684kHz by tuning C1 and C2, which is more than sufficient for it to be used for generating either sideband. L1 was wound with 36swg enamelled wire, on a ¼in diameter former with an iron-dust core. To ensure

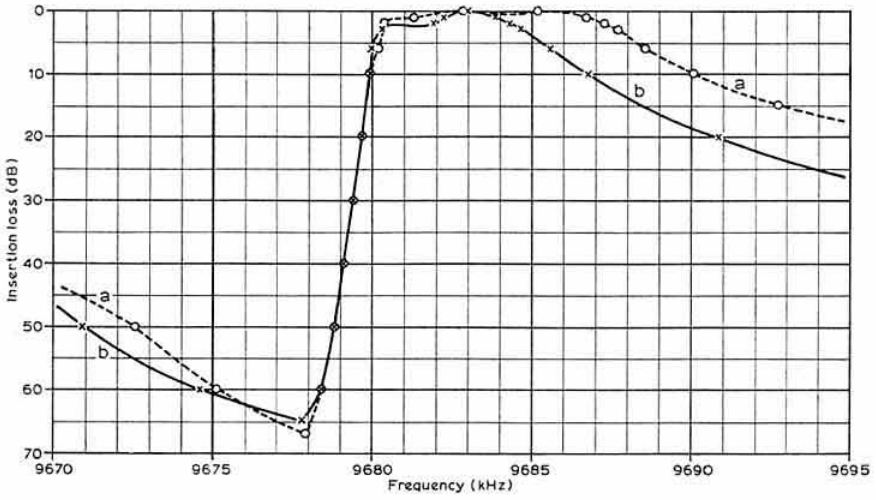


Fig 18. Upper sideband filter. (a) Without C6, 7, 8; (b) With C6, 7, 8

oscillation the dots indicating the start of each winding must be observed.

Conclusion

If a single outstanding factor were to be chosen from all the measurements which have been made it would be that it is almost impossible not to produce a filter of some sort, if the guidelines which have been given are followed. This should be sufficient to encourage even the most hesitant to try to construct their own crystal filters.

Another remarkable feature is the apparent lack of attention this type of filter has received, both in amateur and professional journals, and the author would be pleased to be advised of any relevant publications known to readers.

Results of readers own experiments would also be received with great interest.

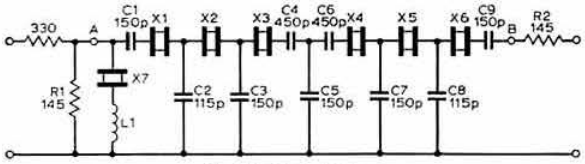


Fig 19. Dishal filter.
X7 and L1 are an optional addition to produce a symmetrical filter. L1 is 10 turns 28swg enam on ¼in diam iron-dust cored former. X1-X7 9,681·2kHz.
C1, C3, C5, C7, C9: see text, capacitance depends on bandwidth required. C2, C8 = 0·75C1

$$C4, 6 = \frac{1}{\frac{1}{C2} - \frac{1}{C5}} \qquad R1, 2 = \frac{1}{2\pi f C2}$$

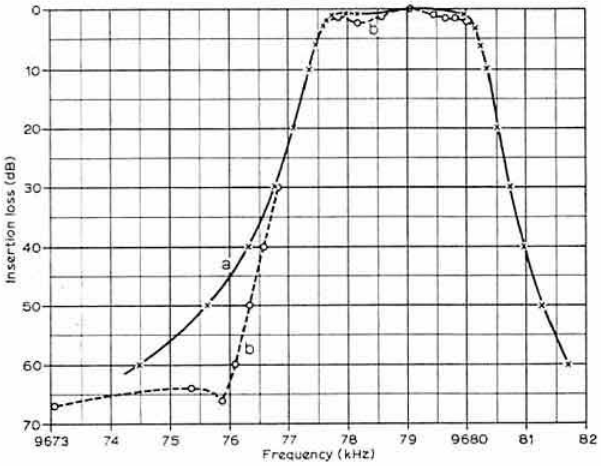


Fig 20. Dishal filter. (a) Six-crystal filter; (b) Seven-crystal filter

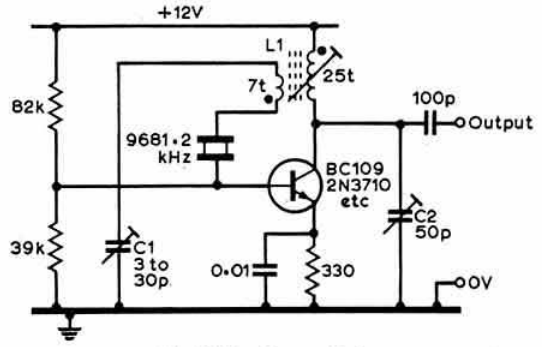


Fig 21. Carrier oscillator

References

[1] "Technical Topics", Pat Hawker, G3VA, *Radio Communication* September 1976.

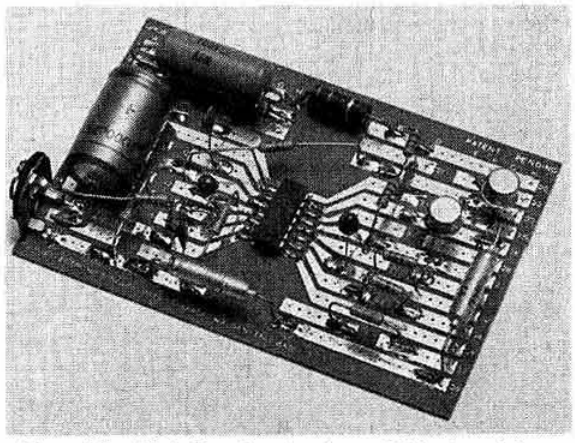
[2] "Modern network theory of single sideband crystal ladder filters", Milton Dishal, *Proceedings IEEE* Vol 53, No 9, September 1965, pp1205-1216.

[3] *Electronics Engineers Handbook*, Fink(Ed), McGraw Hill, Tables 12-1 to 12-7. Section 12, Filters, coupling networks and attenuators, by Milton Dishal. □

NEW PRODUCT

"Blob Board" prototype pcb

The manufacturer of the S-DeC board has introduced a new type of pcb, the Blob Board, which, it is claimed, will permit the user to go from circuit diagram to completed board in a short time and without the use of chemicals. The component layout can be drawn directly on the Blob Board with an ordinary felt pen, and the components are then laid out with leads on copper strip. A few blobs of solder and the circuit is complete. Components can be re-used and re-soldered, thus facilitating amendments. All Blob Boards are roller tinned. They are designed for the home constructor and prototype engineer and are said to be normally half the price of competitive boards. There are 12 different Blob Boards which, between them, will accommodate all components from discretes to ICs.



One of the Blob Boards, showing a finished prototype

They are obtainable from P.B. Electronics (Scotland) Ltd, 57 High St, Saffron Walden, Essex CB10 1AA, from whom leaflets and further information can be obtained by sending an sae.

Some experiments with high-frequency ladder crystal filters

by J. A. HARDCASTLE, G3JIR*

Part 4. SSB transmitter filters

Feedback

Since the publication of the first three articles on ladder crystal filters [1-3] correspondence [4] and published suggestions [5] have pinpointed a need for a cheap ssb transmitter filter using crystals which are not necessarily identical in frequency. These notes describe briefly two filters which use crystals removed from an ITT 445LQU901A filter.

445LQU901A 50kHz channel filters

These filters contain eight crystals whose reference numbers indicate that there are four pairs of nominally identical crystals. As shown in Table 1 they fall into two groups of four frequencies which are sufficiently similar to allow them to be used in a four-crystal ssb filter.

SSB filters

Figs 16 and 17 [3] showed how a high performance could be obtained from a few crystals if all the selectivity were concentrated on one sideband. They were adjusted to provide uniformly high selectivity over a wide stopband and it has been pointed out by John Haine, ex-G8CEG, [4] that the rate of cut-off can be further increased by tuning all the crystals to the same peak loss frequency. Fig 22 shows the circuits which were used. They are slightly different from Figs 16 and 17 because of the need to trim each crystal to the peak frequency. C1 to C4 are 40pF compression trimmers although the actual capacitance setting is noted by each trimmer because it reflects the amount by which each crystal's frequency was shifted and may be helpful to constructors.

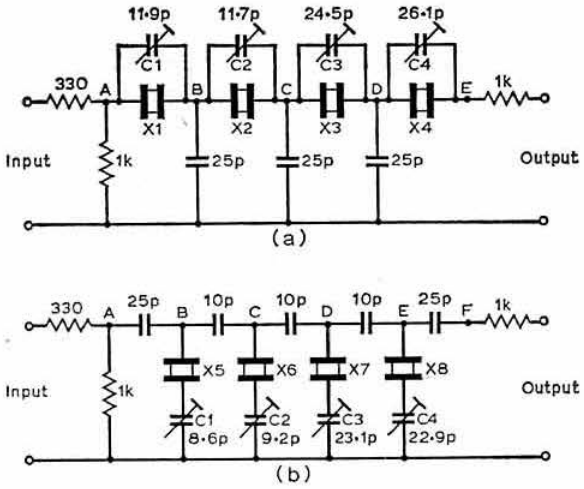


Fig 22. SSB filters. (a) Lower sideband. (b) Upper sideband

Tuning

The filters are tested using the test set of Fig 12[2]. During tuning each crystal must be isolated from its neighbours, and upper and lower sideband filters each need a slightly different procedure to be followed.

In Fig 22(a) the first crystal is isolated by linking points B and E. The parallel resonant frequency (ie frequency of minimum output) of XI is then moved closer to the passband by increasing the capacitance of C1. To tune the second section, point A is linked to B, point C to E, and C2 is trimmed to tune X2 to the same frequency as X1. The same procedure is applied to the remaining sections to complete the tuning.

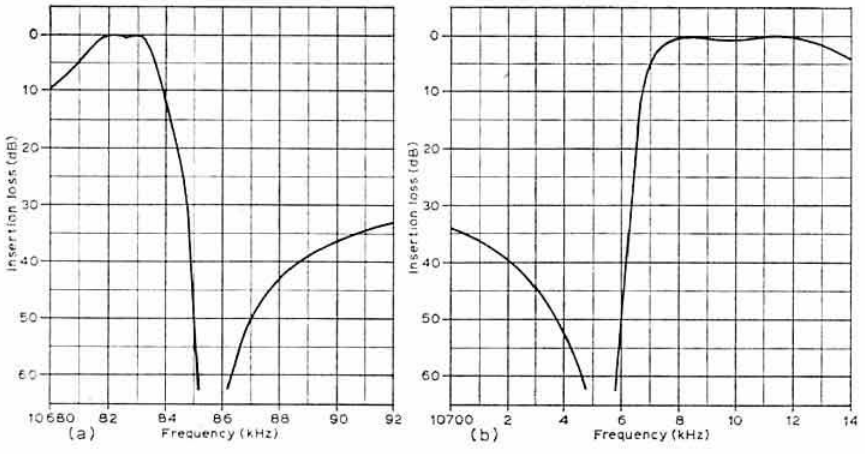


Fig 23. SSB filter frequency response. (a) Lower sideband (b) Upper sideband

*82 Acacia Avenue, Huyton, Liverpool L365TP.

Table 1

Crystal case reference	Fig 22 reference	Series resonant frequency (kHz)	Parallel resonant frequency (kHz)
0074	X1	10,680·7	10,699·9
0074	X2	10,680·9	10,699·6
0075	X3	10,683·3	10,702·5
0075	X4	10,683·3	10,702·1
0076	X5	10,700·6	10,720·0
0076	X6	10,700·3	10,719·9
0077	X7	10,702·8	10,722·8
0077	X8	10,703·0	10,722·7

In the upper sideband filter, Fig 22(b), the peak rejection frequency is determined by the series resonance of the crystals. To tune the first section, A is linked to B, and B to F. In this case *reducing* the capacitance of C1 will move the series resonance of X5 nearer to the passband so it is useful to set all the trimmers to maximum capacitance initially. For the second section, A is linked to C, and C to F, and similarly for the remaining sections.

Performance

Fig 23 shows the frequency response of each filter. The choice of the best maximum rejection frequency will probably require several attempts at tuning. The closer the notch is positioned to the passband the steeper the rate of cut-off; however, a point is reached when the passband becomes restricted and this is illustrated in Fig 23(a). This filter would give a better overall response if the rejection notch were moved slightly higher in frequency.

A transmitter incorporating these filters requires a lowpass filter to limit the audio bandwidth before it is applied to the modulator. Without this additional filter the bandwidth of the transmitted signal will be unnecessarily high.

Conclusion

These additional designs have been reported to show that the cost of filters need no longer be a barrier to home construction of ssb transmitters, even if new crystals have to be purchased at commercial prices.

References

[1] "Some experiments with high frequency ladder crystal filters", J. A. Hardcastle, G3JIR.
Part 1, *Radio Communication* December 1976.
[2] Part 2, January 1977.
[3] Part 3, February 1977.
[4] Private communication, 17 January 1977.
[5] "Technical Topics", *Radio Communication* June 1977. □

Calculation of distances from QTH Locator codes or latitude and longitude using scientific calculators

by J. H. BOWEN, CEng, MIERE, G8DET;* R. C. HARVEY, TEng (CEI), MITE, G4BBR; and A. J. HARRIES

A NUMBER of methods have been described for calculating the distance between two stations whose positions are defined by QTH Locator codes. Some methods make use of computers while others use programmable calculators, both being very efficient, especially when a large number of calculations are required as in tallying up the score for a contest. The authors feel, however, that there is also a need for a method of calculating these distances using relatively cheap scientific calculators, in fact one machine currently available at under £10 does the job admirably.

The described process is in two main parts, the first being the conversion of the QTH Locator code into degrees longitude east and degrees latitude north. Locations to the west of Greenwich are expressed as negative degrees east. The conversion is easily accomplished with the help of Table 1 which assumes the QTH is in the centre of the QTH Locator code tertiary "square".

1. G4BBR QTH, QTH Locator code YL10e.
   Longitude 1 (−4) + (1·90) = −2·10°E
   Latitude 1 (51) + (0·90) = 51·90°N
2. G8DET QTH, QTH Locator code AL23b
   Longitude 2 (0) + (0·57) = 0·57°E
   Latitude 2 (51) + (0·73) = 51·73°N

Due to the fact that the QTH Locator code system does not define a unique location, stations have been worked on 144MHz with the same QTH Locator codes separated by 26° of latitude, ie 2,889km. Reference to a map will show if the correct latitude (or longitude) has been chosen. A maximum error of ±7km is possible between two European stations due to the relatively large area covered by the QTH Locator code tertiary "square". The error may be halved if the actual latitude and longitude, expressed in decimal degrees, is used for the home QTH. For dx record purposes the latitude and longitude for both stations should really be used together with a multiplier of 111·12.

The second stage involves the calculation of the distance between the defined QTHs, and the standard formula for this is given at the foot of this page.†

A number of calculators have been tried on the above formula and all those listed below will work out the distance without having to resort to pen and paper during the calculation:

Hewlett Packard Scientific range;

*62 St Clere's Way, Danbury, Chelmsford, Essex.

† Distance = $\cos^{-1}$ [Sin(Lat 1) × Sin (Lat 2) + Cos (Lat 1) × Cos (Lat 2) × Cos (Long 1−Long 2)] × 111km
where Long 1 and Lat 1 is the location of the home QTH and Long 2 and Lat 2 is the location of the distant QTH.